In the Specification

Please amend the specification of this application as follows:

Rewrite the paragraph at page 2, lines 6 to 6 as follows:

--where ~~fd~~ $\underline{f_d}$ is the frequency deviation and assumed to be 75 KHz for commercial FM. ~~fd~~ $\underline{f_d}$ represents the maximum shift of f(t) relative to the carrier frequency.--

Rewrite the paragraph at page 2, lines 10 to 11 as follows:

--2) ~~fm(t)~~ fm(t) = BBMUX = [Left (t) + Right (t)] + Ap Sin (ω_pt) + [Left(t) - Right(t)] Sin (2ω_pt)--

Rewrite the paragraph at page 2, line 13 as follows:

--~~fm(t)~~ fm(t) = the time varying value of the composite signal--

Rewrite the paragraph at page 3, lines 10 to 10 as follows:

--As is known, an FM signal at baseband, or zero frequency, is the result of "mixing out" the carrier frequency from the FM signal, as shown in Figure 1. Thus, the carrier frequency no longer appears in the FM equation. The generation of complex data is the result of mixing FM signal with a cosine and sine local oscillator (LO), and the process can be derived mathematically by using the complex version of the FM equation. The mixing process is a simple multiplication of signals where the cosine mixing term and sine mixing term are multiplied with the incoming FM signal. For baseband results, both mixers oscillate at the FM carrier frequency ω_c. As shown, the total mixing operation produces a real (in-phase) and imaginary ~~(quadrature-phase~~ (quadrature-phase) baseband component.--

Rewrite the paragraph at page 3, line 21 as follows:

--1) $x_{FM}(t) = A_c \cos (2\pi f_c \cancel{+ \emptyset(t)_{FM},}\ \underline{\emptyset(t)_{FM}),}$ and since $\omega_C = 2\pi f_c$--

Rewrite the paragraph at page 3, line 22 as follows:

--1A) $x_{FM}(t) = A_c \cos (\omega_c + \cancel{\emptyset(t)}\ \underline{\emptyset(t))}$--

Rewrite the paragraph at page 4, line 3 as follows:

--~~5.~~ <u>5)</u> $\cancel{x_{FM}^{(t)}}\ \underline{x_{FM}(t)} = \mathrm{Re}\{ A_c e^{j[\omega_c t + \emptyset(t)]}\}$--

Rewrite the paragraph at page 4, lines 4 to 6 as follows:

--Thus, it is seen that the input $\cancel{X_{FM}t}$ $\underline{X_{FM}(t)}$ on line 10 of Figure 1 can be expressed by an equation in the form of Equation 1A. Consequently, the in-phase component will be $X_{FM}(t) \cos (\omega_c t)$ and is represented by:--

Rewrite the paragraph at page 4, lines 7 to 8 follows:

--4) in-phase = $\cancel{\mathrm{Re}[A_e e^{j[\omega_e(t)+\emptyset(t)]} \cos\ \omega_e t}$ $\underline{\mathrm{Re}[A_c e^{j[\omega_c t+\emptyset(t)]}]\cos(\omega_c t)}$ which as will be appreciated by those skilled in the art can be reduced to:--

Rewrite the paragraph at page 4, line 11 as follows:

--6) Q-phase = $\cancel{\mathrm{Re}[A_e e^{j[\omega_e t+\emptyset(t)]}]\sin(\omega_e(t))}$ $\underline{\mathrm{Re}[A_c e^{j[\omega_c t+\emptyset(t)]}]\sin(\omega_c t)}$ which is reduced to--

Rewrite the paragraph at page 4, lines 16 to 22 as follows:

--Referring now to Figure ~~2,~~ <u>3,</u> there is shown the complex Cartesian coordinate system and the complex unit circle $|z|=1$. From Equation 8, the complex equation for an FM signal at baseband, a complex-valued FM sample can be represented by a vector on the complex unit circle having an amplitude and a phase angle. A complex sample gives two pieces of information, a real and

imaginary component. The polar form of a complex number z, where ~~x=x+jy,~~ z=x+jy, can be represented by the following equations:--

Rewrite the paragraph at page 5, line 2 as follows:

--10) $\cancel{r=\sqrt{x^2+y^2}}$ $r=\sqrt{x^2+y^2}$ --

Rewrite the paragraph at page 5, lines 19 to 24 as follows:

--Equation 12 is used to directly relate the phase difference between two complex-valued samples. In the previous example, suppose the sinusoid is sampled at a rate eight times greater than its frequency. Applying Equation 12), each vector will travel ~~(1/8(360°)~~ (1/8(360°)) = 45° from the previous sample's location. Furthermore, at the Nyquist sample rate, or $2f_{max}$, each successive vector will travel 180° from the last vector's position.--

Rewrite the paragraph at page 6, lines 1 to 20 as follows:

--Thus, it is seen that this FM demodulation technique requires division and arctan calculations. That is, $\theta = \arctan x/y$, where X = IM and Y = RL so that $\cancel{\theta = \arctan IM/RL}$ $\theta = \arctan(IM/RL)$. However, as is well understood by those skilled in the art, typical fixed point DSPs (FP DSP) do not include dedicated hardware for such calculations. Consequently, other various methods must be used to accomplish these calculations in the commercial or typical DSP, such as for example TI's fixed point DSP TM5320C6201. The Newton-Raphson method is one of the most commonly used methods for division calculations for FM demodulations that require the division of IM/RL (IM = imaginary/ RL = real). This technique gives the maximum accuracy of the calculation requiring the use of a reciprocal instruction. Another way of performing division in a fixed point DSP is by using "conditional subtraction" method. A

third method and the fastest of all is the use of look-up tables for 1/RL and then multiplying by IM (the imaginary term). Unfortunately, there is a disadvantage of using look-up table methods because of the very small values of RL encountered. If the term RL = cosine Θ is used, the size of the look-up table must be very large for acceptable accuracy. The obvious disadvantage of such a large look-up table is the inefficient use of memory. The method of the present invention, however, uses the fast look-up table method with an approximation. This approximation overcomes the memory problems faced by the conventional look-up table approach.--

Rewrite the paragraph at page 7, lines 2 to 23 as follows:

--As will be appreciated by those skilled in the art and discussed above, one of the calculations encountered in the demodulation and decoding of digital complex base-band FM signals (BBMUX) is dividing the imaginary term IM by the real term RL. That is, IM/RL. The inefficient memory usage which is a result of the extensive look-up table required by the very small values is overcome by the teachings of the present invention by utilizing the fact that the real term RL is an in-phase representation, and the fact that the real term RL as used in a complex FM base-band system is determined from the function [cosine Θ]. As is well known, the solution of the cosine function will always, of course, range between -1 and +1 for all values of Θ. Therefore, by adding the number two (2) to the RL cosine Θ value, the range of the denominator will be shifted from between -1 and +1 for 1/RL to between 1 and 3 for ~~1/RL+2~~ <u>1/(RL+2)</u>. That is, 2 - 1 = 1, and 2 + 1 = 3. Thus, the addition of two (2) to cosine Θ solves the problem of having a very large look-up table required by the small values of RL as they pass through 0. As will become clear hereinafter, addition of two (2) to the RL value results in a reduction of

amplitude of the demodulated signal by a constant factor K that is substantially uniform over the full range. Therefore, if the output of the demodulation process is multiplied by this same constant factor K, a demodulation signal can be obtained which has an excellent signal to noise ratio. That is, the resulting BBMUX signal is almost identical to that calculated by the Newton-Raphson method and perhaps superior to the "conditional subtraction."--

Rewrite the paragraph at page 8, lines 20 to 23 as follows:

--Figure 5 illustrates the magnitude of the BBMUX signal as determined if Newton-Raphson was used to calculate the value of 1/RL and using the look-up table of the present invention to calculate ~~1/RL + 2~~ 1/(RL+2).--

Rewrite the paragraph at page 9, line 15 as follows:

--Assuming F_P = 19 ~~KH_z,~~ KHz, then ~~WP~~ ω_p = 2 x π x 19 KH_z = 119.38 ~~KH_z~~ KHz--

Rewrite the paragraph at page 11, line 13 to page 12, line 19 as follows:

--Therefore, according to the teachings of this invention, the fact that the RL value is an in-phase representation allows the RL part or value to be manipulated in such a way so as to solve this difficult problem of having a look-up table for very small values. More specifically, and as shown in step 34 of Figure ~~2~~ 2A, by adding a numeral value N to the value of RL (which it will be appreciated, RL ranges from -1 to +1) before calculating the reciprocal value ~~(i.e. 1/RL + N)~~ (i.e. 1/(RL+N)) will provide a significant change from the original value of 1/RL. For example, as mentioned above, since RL varied between -1 to +1, the reciprocal value of 1/RL also ranges from ~~-1 to +1,~~ (-∞,-1] and [1,+∞), and therefore, the look-up table must include values that

approach and pass through zero. However, if the number two (2) is added to RL prior to taking the reciprocal value, the final value of ~~1/RL+2~~ 1/(RL+2) will range from 1 to 1/3 and does not go through zero. Therefore, it is seen that there are no values less than 1/3 and no values greater than 1. Therefore, the size of the look-up table indicated at step 36 of Figure 2A may be significantly reduced since the values do not pass through zero and there are no extremely small values. It will also be appreciated that the value of N in the equation ~~1/RL + N~~ 1/(RL+N) is preferably chosen to be the integer 2 as this provides the simplest and most effective results. However, it should be understood that other values of N may be used advantageously, but if the value of N varies much from the value of 2, the range between the highest value to the lowest value begins to increase significantly thereby again increasing the size of the look-up table. For example, if the value of one (1) was added rather than two (2), the denominator would again go to 0 and consequently the value would increase to infinity, which would make the computations just as difficult as if the number one (1) had not been added. Likewise, if values much greater than 2 are added to RL in the 1/RL function, the separation between the maximum and minimum values decreases. This means that smaller and smaller numbers must be used to distinguish between different values. Consequently, it is believed the practical range of numbers to be added to RL for use with this technique ranges from about a value of RL+1.1 to achieve a total or overall range of between 10 and 0.9 and adding a value of 5 to the value of RL to achieve a total or overall range of between 0.3 and 0.2.--

Rewrite the paragraph at page 12, line 20 to page 13, line 2 as follows:

--Thus, by adding a value N to the function 1/ RL to obtain ~~1/RL+N,~~ 1/(RL+N), the range is substantially shifted and if N is

chosen to be a value between about 1.1 and 5, the problem of having a look-up table with very small values will be solved. After the reciprocal of the term ~~1/RL + N~~ 1/(RL+N) is determined, the arctangent of the resulting value is then determined as indicated at step 38 of Figure 2A. That arctangent value is preferably also determined by referring to a look-up table. --

Rewrite the paragraph at page 13, lines 3 to 8 as follows:

--Further, referring to Figures 5 and 6, assuming the value of N is chosen to be 2, there is an overall reduction of amplitude of the demodulated signal by a factor K that is substantially uniform. In the example illustrated in Figures 5 and 6, the factor K was simply determined by taking the ~~ration~~ ratio of the amplitude of curve 50 and curve 52 at peak 53. That is, $K = \frac{Max\,(idealBBMUX\;\;)}{Max\,(scaledBBMU\;\;X)}$. This ratio was calculated to be approximately 7.--

Rewrite the paragraph at page 13, line 9 to page 14, line 6 as follows:

--Referring again to Figure 5 and as described above, curve 50 represents the magnitude of BBMUX determined by solving for the arctan of IM/RL when computed in the typical and prior art techniques of using Newton-Raphson, conditional subtraction, or a very large look-up table for 1/RL. The curve 52, which has an amplitude that varies to a much smaller degree around the zero base line, but has the same shape as curve 50. The curve 52 is the magnitude of BBMUX determined by solving for the arctan of ~~IM/RL+2.~~ IM/(RL+2). Now, referring to Figure 6, we have the same curve 50 determined by solving the arctangent of X/Y (IM/RL) as displayed in Figure 5. However, the BBMUX curve determined from the arctan of ~~IM/RL + 2~~ IM/(RL+2) has been multiplied by the factor K as indicated by step 40 of Figure 2A and the shape of the curve 54

using ~~(K)(arctan(1/RL+2))~~ (K)arctan(1/(RL+2)) is seen to be almost identical to the curve 50. Therefore, as seen in the flowchart, we now have determined and demodulated the magnitude of the conventional FM stereo signal. Thus, by taking the arctangent of ~~IM/RL+2~~ IM/(RL+2) and then multiplying with the amplification factor K at the output, an excellent approximation to the demodulated signal is obtained. Further, since the look-up table does not go through zero and consequently does not have very small values, memory is saved and the number of cycles required to complete the calculation is reduced considerably which, of course, increases the speed of the computations. After the signal has been demodulated and the BBMUX signal obtained, it is typically filtered as indicated at step 42 of Figure 2A prior to decoding the signal to determine the left and right input signals.--

Rewrite the paragraph at page 15, line 10 as follows:

--12) $$I=\sum_{n=0}^{N-1} fm(n)\cdot\cos(2\pi nF_p/F_s)=\cancel{\sum_{n=0}^{N-1} fm(n)\cdot\cos(\pi n/4}\ \underline{\sum_{n=0}^{N-1} fm(n)\cdot\cos(\pi n/4)}$$ --

Rewrite the paragraph at page 15, line 12 as follows:

--14) $$A_p=(\tfrac{2}{N})\cdot\cancel{\sqrt{I^2+Q^2}}\ \underline{\sqrt{I^2+Q^2}}$$ --

Rewrite the paragraph at page 16, lines 3 to 19 as follows:

--Factors used in calculating the DFT are determined by calculating the $\cos(\pi n/4)$ and $\sin(\pi n/4)$ terms, letter n be integers in the range [0,…7,]. Therefore, only 8 cosine and 8 sine terms are necessary to compute the DFT. Using the non-coherent scheme will require a DFT ~~(differential~~ (discrete fourier transform) snapshot as indicated at block 60 in Figure 7 for 128 points at the pilot signal frequency (as shown at step 46 in Figure 2B) in order to determine the pilot signal magnitude and phase. Since the pilot

frequency was chosen to be 19 KHz in this embodiment, the sampling rate is preferably chosen to be 152 KHz as that number is eight times the pilot frequency. Further, this rate is also above the Nyquist rate of 106 KHz. Therefore, for a time period of 2 π of the signal, the pilot frequency computations need to be performed at 45°, 90°, 135°, 180°, 225°, 270°, 315° and ~~360.°~~ 360°. The prior art method of accomplishing this is to provide a cosine and sine table, each with eight entries of the value of cosine and sine at these eight phase values (8+8 = 16) and then multiplying the resulting value with the input for every eight points for a total of 128 points (8 x 16 = 128). Consequently, as will be appreciated by those skilled in the art, this prior art method requires eight multiplication and eight addition steps.--

Rewrite the paragraph at page 17, line 15 to page 18, line 3 as follows:

--After determination of 128 snapshot points for the DFT ~~(differential~~ (discrete fourier transform) points, the magnitude and phase of the pilot signal is determined as shown at 64 and 66 of Figure 7 and step 49 in Figure 2B. Magnitude and phase of the pilot signal is determined according to techniques well known by those skilled in the art, and as mentioned above the pilot signal phase is provided to the filter bank selector 63 that selects the appropriate polyphase filter from filter bank 62. Of course, once the pilot signal magnitude and phase is known, the demodulated information or message signal can be determined by removing the pilot signal from the poly phase filtered signal received from poly phase filter 62 as shown at 68 in Figure 7 and block 51 in Figure 2B. Then, after filtering, channel separation is accomplished by techniques already known by one skilled in the art as indicated at step 53 and block 70 of Figure 2B and 7 respectively.--

Rewrite the paragraph at page 19, lines 11 to 16 as follows:

--As seen, the number of cycles ranges from 346 for the prior art technique ~~and~~ to 83 cycles for the present technique. Therefore, it is seen that using the techniques of this invention to perform division using a look-up table with an approximation and implementation of a faster DFT snapshot for obtaining samples, a fixed point DSP may readily be used for FM demodulation and decoding for a digital radio application which operates in a fast and efficient method.--

Rewrite the paragraph at page 19, line 16 to page 20, line 4 as follows:

--Therefore, there has been described to this point a process for demodulating and decoding a conventional stereo FM signal using a look-up table for taking the reciprocal of RL (i.e. ~~1/RL+2~~ 1/(RL+2) rather than 1/RL). When the resulting reciprocal value is determined, a very accurate approximation to a transmitted signal is recovered by multiplying with a factor K. A method for reducing the number of computation steps in a non-coherent technique for decoding the left and right channels has also been described. However, although the invention has been described with respect to specific methods, it is not intended that such specific references be considered limitations upon the scope of the invention except as set forth in the following claims.--